BANDPASS PREDISTORTION METHOD AND APPARATUS FOR RADIO TRANSMISSION

CROSS-REFERENCE TO RELATED APPLICATION

[0001] This application is related to United States Patent Application Serial No. 09/624,149 filed July 24, 2000.

FIELD OF THE INVENTION

[0002] The present invention pertains to an apparatus for and a method of applying both amplitude predistortion and phase predistortion to a modulated baseband signal. More particularly, the present invention pertains to an apparatus for and a method of generating an amplitude modulated radio frequency signal by amplitude predistorting its baseband signal, using the inverse hyperbolic tangent of a value based on the envelope of the baseband in-phase and quadrature components, and phase predistorting the baseband signal, using the hyperbolic tangent of that value.

BACKGROUND OF THE INVENTION

[0003] Environments such as commercial airliners frequently have several radios that operate at different frequencies. Not only must these radios avoid interference with each

other, but also they must meet spectrum mask requirements imposed by regulatory agencies, such as the United States Federal Communications Commission. The output from the solid state power amplifier of such a radio often includes distortion that can be characterized by a hyperbolic tangent function. Both amplitude distortion and phase distortion may occur. The transmit spectrum of such a radio signal can spread near the desired signal band if the envelope of the transmitted signal is not constant, particularly if the transmitter power amplifier is being driven into soft saturation. While spurious emissions might be reduced by predistorting of the radio frequency signal envelope just before transmission to the output power amplifier, this requires analog multipliers. Even then, if noise is picked up in the multiplier circuit, that noise will modulate the desired signal and pass through to the output.

[0004] One approach to overcoming power amplifier nonlinearity utilizes the function $f(x) = 2x/(1+x^2)$ for amplitude predistortion and the function $ph(x) = (Bf(x))/6 = 2Bx/6(1+x^2)$ for phase predistortion, where x is the instantaneous value of the envelope. Another approach to overcoming amplitude distortion is to utilize the "cuber" function $f(x) = x+x^3/3$, where again x is the instantaneous value of the envelope. These approaches have been found to provide less than optimum linearity in the power amplifier output.

SUMMARY OF THE INVENTION

[0005] The present invention is an apparatus for and a method of amplitude and phase distorting a modulated radio frequency signal such that after passing of the distorted signal through a non-linear power amplifier, undesirable spurious emissions in the resulting spectrum are reduced. In accordance with the present invention, a complex amplitude modulated baseband signal, having an in-phase component I and a quadrature component Q, is sampled to obtain k samples I_k of the in-phase component and k samples Q_k of the quadrature component, and the magnitude of the envelope of the baseband samples is determined. A distortion factor based on the product of the hyperbolic tangent ("tanh") and the inverse hyperbolic tangent or archyperbolic tangent ("atanh") of a scaled value of the

complex baseband sample magnitude is used to multiply each sample of the in-phase component and of the quadrature component so as to provide predistorted components. These predistorted components are combined and used to provide a distorted radio frequency ("RF") signal which is applied to the power amplifier. The power amplifier distortion cancels the distortion in the radio frequency signal so that the power amplifier provides a substantially undistorted output signal.

[0006] The scaling factor is obtained by combining a portion of the output signal envelope with the undistorted envelope in a feedback circuit. The feedback circuit preferably computes the mean square error between the undistorted envelope and the output signal envelope. Preferably, to assure that the mean square error is computed correctly, both envelopes are normalized. The mean square error is adjusted by a fixed gain control and integrated, and the result used to scale the undistorted envelope prior to determination of the hyperbolic tangent and archyperbolic tangent functions.

[0007] The envelope of the baseband signal is thus subjected to amplitude and phase predistortion prior to upconversion to the radio frequency signal. This avoids impressing pick-up noise on the transmitted envelope. It is possible to do the predistortion prior to intermediate frequency (IF) and RF bandpass filtering of the radio frequency signal since such filtering has a wide bandwidth, allowing the distorted signal spectrum to pass through the power amplifier.

[0008] Preferably, the predistortion apparatus of the present invention is implemented in a gate array, such as a field programmable gate array.

BRIEF DESCRIPTION OF THE DRAWINGS

[0009] These and other aspects and advantages of the present invention are more apparent from the following detailed description and claims, particularly when considered in conjunction with the accompanying drawings in which like parts bear like reference numerals. In the drawings:

[0010] Figure 1 is a block diagram of an apparatus for generating an amplitude and phase predistorted radio frequency signal in accordance with a preferred embodiment of the present invention;

[0011] Figure 2 is a block diagram of one preferred embodiment of a circuit suitable for use in the apparatus of Figure 1;

[0012] Figure 3 is a graph of results from a simulation comparing the present invention with the prior art;

[0013] Figures 4A-4D plot performance in a simulation of the present invention and the prior art; and

[0014] Figures 5A-5D show the output spectra from a simulation of power amplifiers in accordance with the present invention and the prior art.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

[0015] Figure 1 depicts an apparatus for generating an amplitude and phase predistorted radio frequency signal in accordance with a preferred embodiment of the present invention. A signal source 10 provides a complex baseband signal $xe^{j\varphi_k}$, where x is the envelope of the signal and, for example, may be an Edge GSM or a D8PSK signal. The signal includes an in-phase component I and a quadrature component Q that are normalized and sampled at, for example, 10.5 kilosamples per second (KSPS). From source 10, the samples are filtered in filter circuit 12 to produce smooth transitions between phase symbols. The samples I_k of the in-phase component and the samples Q_k of the quadrature component are applied from filter circuit 12 to a calculation circuit 16 which calculates the magnitude of the scaled complex baseband envelope sample, for example by determining the square root of the sum of the squares of the scaled in-phase component sample and the scaled quadrature component sample.

[0016] Figure 2 is a block diagram of one preferred embodiment of a calculation circuit for determining an approximation of the magnitude of each complex sample k of the

baseband signal. In Figure 2 the samples I_k of the in-phase component and the samples Q_k of the quadrature component are applied to a first detection circuit 18 which determines the maximum of these samples by determining for each sample pair whether the I_k sample or the Q_k sample is the larger. The I_k and the Q_k samples are also applied to a second detection circuit 20 which determines the minimum of these samples by determining for each sample pair whether the I_k sample or the Q_k sample is the smaller. The detected maximum value ("max_k") and the detected minimum value ("min_k") for each sample pair are applied to calculating circuit 22 which computes the value $y_k = ½ (min_k/max_k)^2$.

[0017] The y_k output from calculating circuit 22 is applied as an input to each of five multiplier circuits 24, 26, 28, 30 and 32. The y_k output is also applied to a second input of multiplier 24. As a consequence, multiplier 24 provides as an output the value y_k^2. This y_k^2 output from multiplier 24 is applied to the second input of multiplier 26 and to a negative input to summation circuit 34. The output of multiplier 26 is thus the value y_k^3. This output is applied to the second input of multiplier 28 and to a positive input of summation circuit 34. Multiplier 28 accordingly provides the output y_k^4 which is used as the second input to multiplier 30 and which is applied to a negative input to summation circuit 34. Multiplier 30 then provides the output y_k^5 to the second input of multiplier 32 and to a positive input to summation circuit 34. Multiplier 32 provides the output y^6 to a negative input to summation circuit 34.

[0018] Summation circuit 34 divides the sum of its inputs by 2, thus providing as its output the value $½ (-y_k^2 + y_k^3 - y_k^4 + y_k^5 - y_k^6)$. This signal is applied as an input to summation circuit 36, which also receives as inputs the y_k signal from calculation circuit 22 and the constant 1. The output of summation circuit 36 is thus the value $\{1 + y_k + ½ (-y_k^2 + y_k^3 - y_k^4 + y_k^5 - y_k^6)\}$. This is equal to the value $\{(1 + y_k)/2 + ½ (1 + y_k - y_k^2 + y_k^3 - y_k^4 + y_k^5 - y_k^6)\}$. This signal is applied from summation circuit 36 to one input of multiplier 38, which receives the max_k signal from detection circuit 18 at its second input. Consequently, the output of multiplier 38 is

$(\max_k) \times \{(1 + y_k)/2 + \frac{1}{2}(1 + y_k - y_k^2 + y_k^3 - y_k^4 + y_k^5 - y_k^6)\}$ which is an approximation of $(I_k^2 + Q_k^2)^{1/2}$ and thus an approximation of the magnitude x_k of the sample k.

[0019] The output from the apparatus of Figure 1 is provided by power amplifier 64 to antenna 66. Radio frequency coupler 70 couples a portion of that output to envelope detector 72. The detected envelope is applied to analog-to-digital converter 73 which samples at a high sampling rate, shown in Figure 1 as a sampling rate of 50 megasamples per second (MSPS). The output of analog-to-digital converter 73 is normalized by normalizing circuit 74 so that its maximum valve equals 1. The output of calculation circuit 16 is applied through delay circuit 76 to a positive input of summing circuit 78, while the output from normalizing circuit 74 is applied to a negative input of the summing circuit. The input to summing circuit 78 from calculation circuit 16 represents the envelope before distortion, while the input to summing circuit 78 from normalizing circuit 74 represents the envelope after distortion. Delay circuit 76 assures that each undistorted sample is summed with the normalized output resulting from that same sample. The resulting signal from summing circuit 78 is applied to one input of multiplier 80 which receives a weighting factor of - 8 at its second input. The output from multiplier 80 is applied to one input of multiplying circuit 82 which receives the output from normalizing circuit 74 at its second input. The output from multiplying circuit 82 is applied through low pass filter 84 to sampler 86 which applies samples of that output at periodic intervals of, for example, one minute to integrator 88. The output of integrator 88 is a scaling factor C and is applied to one input of multiplying circuit 90 which receives the x_k outputs from calculation circuit 16 at its second input. The output of multiplier circuit 90 is thus Cx_k.

[0020] The Cx_k output from multiplier circuit 90 is applied as an input to calculation circuit 40 which determines the value of $(\operatorname{atanh}(Cx_k))/Cx_k$. By way of an example, calculation circuit 40 might be a lookup table having values to 16 bits for determining a value $x_k^2/3 + x_k^4/5 + x_k^6/7 + \ldots$ which is an approximation of the value $\{(\operatorname{atanh}(x_k))/x_k\} - 1$. The output of the lookup table then is applied to one input of a summation circuit which

receives the constant 1 at its second input so as to provide an approximation of (atanh $(x_k))/x_k$. It is preferred that calculation circuit 40, when in the form of a lookup table, compute the value of the segment $\{(\text{atanh}\,(x_k))/x_k\}$ -1, and that the constant 1 be added by a summation circuit in order to provide the desired accuracy while maintaining the lookup table of a moderate size.

[0021] The x_k output from calculation circuit 16 is also applied as an input to multiplier 92 which receives the value $\pi/6$ at its second input. The Cx_k output from multiplier circuit 90 is applied to calculation circuit 94 which calculates the value $\tanh(Cx_k)$ and applies that value to an input of multiplier 96. Calculation circuit 94 might be a lookup table, for example. The second input of multiplier 96 receives the value $\pi x_k/6$ from multiplier 92. The output of multiplier 96 is thus $(\pi x_k \tanh(Cx_k))/6 = \Phi_k$. This value is applied to lookup table 98 which provides as outputs the values $I_kN = +\cos(\Phi_k)$ and $Q_kN = -\sin(\Phi_k)$. These values are applied to inputs of multiplier pair 100 which receives the output of lookup table 40 at its second input.

[0022] The output of multiplier circuit 100 is thus the distortion factor $\{(\text{atanh}(Cx_k))/Cx_k\}e^{-j\Phi_k} = D_k$. This output is applied to one input of multiplier pair 44. The samples I_k of the in-phase component and the samples Q_k of the quadrature component are also applied to multiplier pair 44. Each sample of the in-phase component and the quadrature component is thus modified by the respective distortion factor D_k, so that the output of multiplier pair 44 is $x_k e^{j\varphi_k}\{(\text{atanh}\,(Cx_k))/Cx_k\}e^{-j\Phi_k} = D_k x_k e^{-j\Phi_k}$. These samples of the modified signal are resampled in resampling circuit 46 at the same sampling rate as in analog-to-digital converter 73, shown in Figure 1 as a resampling rate at 50 MSPS.

[0023] The resampled output from resampling circuit 46 is applied to multiplier pair 48. Signal generator 50 provides an intermediate frequency signal of a frequency less than half the sampling rate of resampling circuit 46, shown in Figure 1 as a frequency of 17 MHz. Sampling circuit 52 samples the sine and cosine outputs from signal generator 50 at the same sampling rate as resampling circuit 46, shown in Figure 1 as a sampling rate of 50 MSPS.

These sampled sine and cosine signals are applied to multiplier pair 48 so that the multiplier pair provides as outputs the intermediate frequency signals $D_k \times I_k$ sin 17 MHz and $D_k \times I_k$ cos 17 MHz. These signals are added in summation circuit 54, and the resulting predistorted, upconverted intermediate frequency signal is applied on line 56 to digital-to-analog converter 58 which samples at the same 50 MSPS rate as resampling circuit 46.

[0024] The output from digital-to-analog converter 58 is applied to band pass filter 60 which is centered at the 17 MHz frequency of signal source 50 and which has a bandwidth sufficient to avoid distortion of the predistorted envelope, for example a bandwidth of 1 MHz. The output from bandpass filter 60 is upconverted to a radio frequency in upconverter 62 and passed through driver amplifier 68 and power amplifier 64 to antenna 66. If desired, a radio frequency attenuator could be utilized, rather than upconverter 62 and driver amplifier 68. Power amplifier 64 has a transfer function C and hyperbolic tangent distortion so that the output of power amplifier 64 is bctanh $(xe^{j\varphi_k} e^{-j\Phi_k} e^{j\Phi_k} \tanh^{-1}(cx))/cx = bcxe^{j\varphi_k}$, where b is the power amplifier gain.

[0025] The feedback circuit of Figure 1 results in the signal C that is applied from integrator 88 to multiplier 90 converging to the current value of the transfer function C of output amplifier 64. It is possible to set the gain of the feedback loop so that it converges in just a few iterations. The value of the feedback gain - 8 which guarantees stable conversion is upper bounded by the mean square value of the feedback envelope after being normalized by circuit 74.

[0026] Predistorting the digital envelope of the baseband signal before upconversion to the radio frequency, followed by digital-to-analog conversion, in accordance with the present invention avoids impressing of analog pickup noise directly on the transmitted envelope, as would occur if the envelope correction were performed on the radio frequency analog signal. Implementation of the present invention does not require significant hardware. It can be accomplished in software or firmware. Implementation on a gate array, such as a field programmable gate array, is convenient.

[0027] Figure 3 is a plot of power amplifier output as a function of signal input for (1) a computer simulated system in accordance with the present invention with the scaling factor C = 0.7, (2) a computer simulated system utilizing the cuber function $f(x) = x + x^3/3$, and (3) a computer simulated system utilizing the functions $f(x) = 2x/(1 + x^2)$ and $ph(x) = 2Bx/6(1+x^2)$, showing the superiority of the present invention.

[0028] Figures 4A-4D are quadrature amplitude modulation plots. Figure 4A plots the computer simulated output of a linear power amplifier. Figure 4B plots the computer simulated output of a non-linear power amplifier with no predistortion, but with hyperbolic tangent nonlinearity in phase and amplitude. Figure 4C plots the computer simulated output of such a nonlinear power amplifier with predistortion based on the cuber function $f(x) = x + x^3/3$. Figure 4D plots the computer simulated output of such a nonlinear power amplifier with predistortion in accordance with the present invention. As can be seen, the plot for the present invention in Figure 4D is substantially the same as the plot for a linear power amplifier in Figure 4A, while the plots of Figures 4B and 4C are not, again showing the superiority of the present invention.

[0029] Figure 5A shows the computer simulated output spectrum of a linear power amplifier. Figure 5B is the computer simulated output spectrum of a nonlinear power amplifier. Figure 5C is the computer simulated output spectrum of such a nonlinear power amplifier with predistortion based on the cuber function $f(x) = x + x^3/3$. Figure 5D is the computer simulated output spectrum of such a nonlinear power amplifier with predistortion in accordance with the present invention with the scaling factor C = 0.7. The simulated output spectrum of the present invention most nearly matches that of a linear power amplifier, once more showing the superiority of the present invention.

[0030] Although the present invention has been described with reference to preferred embodiments, various alterations, rearrangements, and substitutions could be made, and still the result would be within the scope of the invention.